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Time-Frequency Multiplexing (TFM) of Two NTSC Color TV Signals—Simulation Results

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Simple frequency-division multiplexing (FDM) of two television signals onto a microwave radio channel is frequently unsatisfactory because of nonlinearity-induced crosstalk from one picture into the other. With time-frequency multiplexing (TFM), two successive scan lines (or fields) of one picture are frequency multiplexed so that they can be sent in one line (or field) period. During the next time interval, two successive lines (or fields) from the other picture are transmitted, thus avoiding crosstalk between pictures. To reduce the bandwidth required, one of the two simultaneously transmitted lines (or fields) is sent as an analog differential signal. In this paper, we discuss computer simulations of these techniques with application to a 20-MHz-bandwidth microwave radio channel. Effects of filtering and nonlinearities are included insofar as possible.

I. INTRODUCTION

It has been known for some time that frequency-division multiplexing (FDM) of two 4.2-MHz National Television System Committee (NTSC) color television signals onto one microwave radio channel is often unsatisfactory because nonlinear distortion causes visible crosstalk between the two pictures. Also, we suspected that 2:1 time compression followed by time multiplexing would lead to problems because of the color carrier (initially 3.58 MHz) being transmitted at 7.16 MHz, where it is much more susceptible to degradations such as selective fading. However, there are other techniques which show promise of not having such problems.

Here we propose time-frequency multiplexing (TFM), which involves sending two successive scan lines (or fields) from one picture during one line (or field) interval, followed by two lines (fields) from the other picture in the remaining time interval. The two successive lines (fields)

from one picture would be transmitted simultaneously via frequency-division multiplexing. Multiplexing in this manner avoids crosstalk from one picture into the other which would otherwise be caused by nonlinearities in the transmission if simple FDM were used. We hypothesized that nonlinearity induced crosstalk between successive lines (fields) in the same picture will be much less serious because of their high similarity and the fact that the ghost of a picture overlayed upon itself is practically invisible. With these methods the two incoming pictures would have to be in phase synchronism.

While these methods seem feasible on a qualitative basis, quantitatively many questions arise about their behavior. For example, if one of the successive lines (fields) is sent as a differential signal, e.g., line-to-line difference or field-to-field difference, how much can the bandwidth of the differential signal be reduced before detectable picture degradation results? Also, how much intermodulation crosstalk can be tolerated?

To take a first step toward answering these questions and to gain a rudimentary idea as to the limitations of these techniques, we undertook computer simulations. The computer facility display had a resolution of only 256 × 256 pels (picture elements), which is only about one-quarter the resolution of an NTSC picture. Thus, each processed picture can be regarded as one quadrant of a full-resolution NTSC picture. Modulation and filtering inherent in NTSC color multiplexing and demultiplexing were realized by digital filters operating at a simulated sampling rate of four times the color subcarrier frequency. We maintained full NTSC bandwidth for luminance and chrominance, insofar as possible, even though very few home or studio monitors in the U.S. are capable of displaying full resolution (see Appendix A for details). We simulated time and frequency multiplexing inherent in the transmission methods under discussion using baseband digital filters. Effects of nonlinearities were also simulated at baseband, as will be described below.

II. DESCRIPTION OF THE METHOD

Long-haul broadcast color TV transmission is presently by FM, as shown in Fig. 1. One might suggest transmitting a second vestigial-sideband (VSB) signal via FDM as shown in Fig. 2. However, difficulties immediately arise: (*i*) The microwave amplifier nonlinearities can cause intermodulation crosstalk, causing ghosts at an unacceptable level. (*ii*) Modulating the baseband plus VSB signal onto an FM carrier may exceed the bandwidth allowed, since the modulation is not strictly narrowband FM. Carson's rule estimates the required rf bandwidth in Fig. 2 as ≈ 2 (10 MHz + deviation).

Ghosts caused by nonlinearity-induced intermodulation can be made

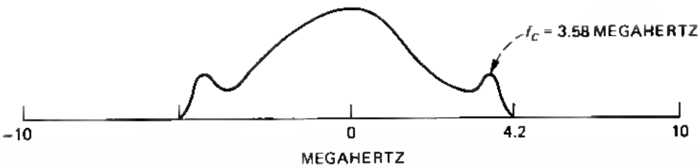

Fig. 1—TV transmission via narrowband FM. The bandwidth actually used is 8.4 MHz plus the bandwidth used by sidelobes (not shown here). If the peak FM deviation is ΔF, then Carson's rule estimates the required bandwidth as 8.4 MHz + $2\Delta F$. The color subcarrier frequency is $f_c \approx 3.58$ MHz.

much less visible if *tim*e-division multiplexing of the two picture signals is employed instead of frequency-division multiplexing. As explained in the introduction, this entails sending two lines (or fields) from one picture in one line (field) interval followed by the same information from the other picture in the next time interval. This requires that the two incoming pictures be synchronized with each other at all times (if this is not the case, a synchronizing device is needed).* Time multiplexing eliminates crosstalk between the two pictures.

Unfortunately, if FDM is used to transmit simultaneously two successive lines (or fields), nonlinearity-induced crosstalk between these two components will still exist. However, two successive lines (or fields) from the same picture tend to be very much alike. Thus, overlaying the ghost of one onto the other should be practically invisible.

From Fig. 2, we see that there may not be enough bandwidth to send two successive lines (or fields) using simple vestigial-sideband in the upper band. However, if a difference signal, e.g., line-to-line difference or field-to-field difference, is transmitted in the upper band, then the aforementioned difficulties can be reduced. A line (or field) difference signal has much less power than the input video and, judging from picture coding experience, probably requires less bandwidth for transmission as well. It has little or no power near DC and, therefore, lends itself to single-sideband modulation (SSB). The arrangement of Fig. 3 may be suitable. In this case the differential signal in SSB modulated, added to the baseband signal and passed to the radio system for transmission via FM. The single-sideband carrier could be transmitted during horizontal blanking so that it would not use up valuable bandwidth during the visible part of the picture, or it could be generated from the color subcarrier.

In deriving the differential signals, differences should be taken

* Synchronization of the two pictures should be such that horizontal sync pulses align. In this way, switching from one picture to the other can occur, for example, following the color burst, and switching transients will not be visible in either picture. An ordinary frame synchronizer should suffice.

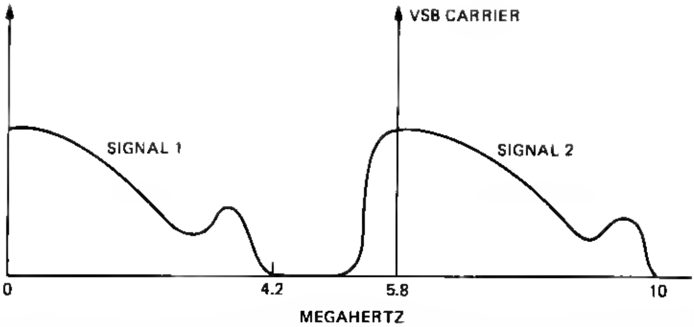

Fig. 2—Transmitting two TV signals via simple frequency-division multiplexing using vestigial-sideband in the upper frequency band. This scheme is unsuitable because of (*i*) nonlinearities and (*ii*) the fact that tbe modulation is not strictly narrowband FM and may require more than the allowable bandwidtb according to Carson's rule, i.e., FM bandwidth ≈ 2 × (message bandwidth + deviation).

between picture elements which on the average have the same color subcarrier phase. This prevents a strong color subcarrier component from appearing in the difference signal. For example, in Fig. 4 three successive lines in one field are shown, and the center one is to be sent via a line difference signal. Pels A, B, C, D, and X all have the same color subcarrier phase (on a flat-colored area). Thus, a suitable line differential signal D_L would be X minus an average of A, B, C, and D. At the receiver, pels A, B, C, and D are averaged in exactly the same way, and the received differential signal is added to recover a replica of the original center line.

In Fig. 5 the center line is from the present field while the other two are from the previous field. Pels A, B, C, and X all have the same color subcarrier phase on a flat-colored area. A suitable differential signal

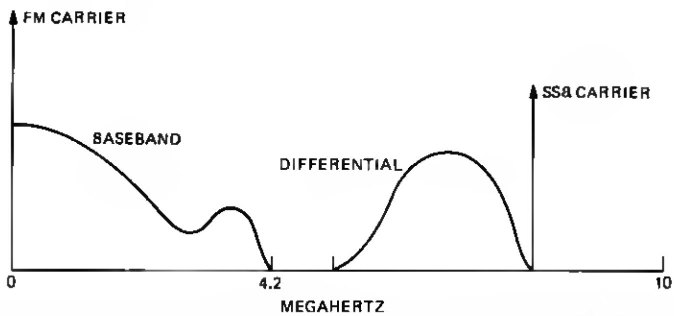

Fig. 3—Time-division multiplexing (baseband *and* differential signals are from the same source). If one of the two lines (or fields) is sent as a differential signal, then it can be placed into the upper band via single-sideband modulation prior to transmission over the radio facility. The single-sideband carrier could be transmitted sometime during horizontal blanking so that it would not use up valuable bandwidth during the visible part of the picture, or it could be generated from the color subcarrier.

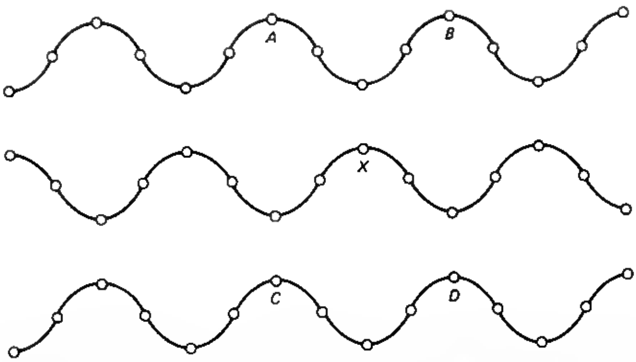

Fig. 4—Three successive lines in one field are shown, and the center one is to be sent via a line-difference signal. Pels A, B, C, D, and X all have the same color subcarrier phase (on a flat colored area). Thus, a suitable line differential signal D_L would be X minus an average of A, B, C, and D, i.e., $D_L = X - \frac{1}{4}(A + B + C + D)$.

D_F is X minus an average of A, B, and C. Since C is closer to X than A or B, it should be weighted relatively more, i.e., $\alpha > \frac{1}{3}$ should be used. At the receiver, we recover the center line by forming the weighted average and adding the received differential signal. In both Figs. 4 and 5 the pel spacing implies a sampling rate of four times the color subcarrier frequency, which is the rate used in the computer simulations.

III. BANDWIDTH REQUIRED FOR DIFFERENTIAL SIGNALS

The first question which we confronted in the computer simulations was "How much can the differential signal be bandlimited before

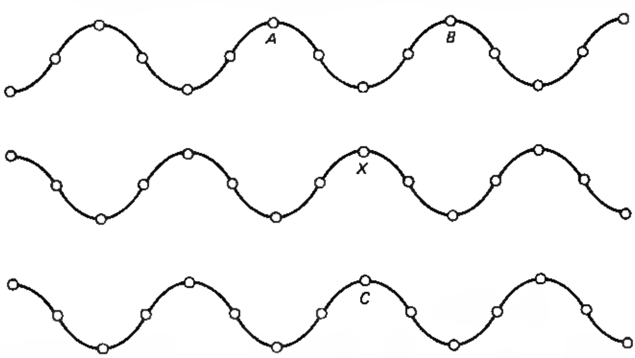

Fig. 5—Three successive lines in one frame are shown, and the center one is to be sent via a field difference signal. The center line is from the present field while the other two are from the previous field. Pels A, B, and C all have the same color subcarrier phase on a flat area. A suitable differential signal D_F is X minus an average of A, B, and C, i.e., $D_F = X - [\alpha C + (A + B)(1 - \alpha)/2]$. Since C is closer to X than A or B, it should be weighted relatively more, i.e., $\alpha > \frac{1}{3}$ should be used. At the receiver the center line is recovered by forming the weighted average and adding the received differential signal.

detectable picture distortion results?" In the simulations, we maintained full NTSC resolution in the color carrier modulation and demodulation (see Appendix A for details), and used a monitor with separate red, blue, green (*RBG*) inputs for display. We used slides from the Society of Motion Picture Television Engineers (SMPTE) with R, B, and G components digitized to eight bits by means of a flying spot scanner for the pictures.

Starting with *RBG* signals for each picture, we first produced NTSC composite signals as described in Appendix A. From these, we obtained reference pictures by NTSC demodulation into the R, B, G components required for display.

To test line-differential transmission, we computed the line difference shown in Fig. 4, from the NTSC composite signals for alternate lines in each field. We then bandlimited the differential signal to various bandwidths, using finite impulse-response digital filters to avoid phase distortions. The bandlimited differential signals were then added to the appropriate averaged signals in alternate lines to recover replicas of the original baseband composite signals. The composite signals were then demodulated to obtain processed pictures.

We compared each processed picture with its corresponding reference picture by switching between the two on the same monitor and closely inspecting small areas for perceivable change. From this we concluded that as long as the line differential signal had a bandwidth of at least 3 MHz, no perceivable difference would result. Below 3 MHz, a slight vertical color shading could be detected in a few areas, such as the boundary beween red lipstick and white teeth.

To test field-differential transmission, we carried out the same procedure using the differential signal shown in Fig. 5. We examined several values of α in the range 0.4 to 0.6, but the results were the same in all cases. We concluded that as long as the field differential signal had a bandwidth of at least 2 MHz, no perceivable difference would result between the reference pictures and the processed pictures. To test the effect on movement rendition, we carried out the processing on sequences of frames moving in pure translation, vertically and horizontally. The results were the same as for still frames and are summarized in Table I.

It is certainly possible to generate test signals which cannot be transmitted without degradation by the methods suggested here. For example, monochrome vertical stripes at the color subcarrier frequency generate a strong line-differential signal at the color subcarrier frequency which would not pass through a 3-MHz low-pass filter. If the vertical interval test signal (VITS) happens to fall on a line that is transmitted via a bandlimited differential signal, then degradation results. The VITS is, after all, designed to test baseband transmission

Table I—Bandwidths required for transmitting differential signals if no perceivable picture degradation is allowed.

Signal	Required Bandwidth
Baseband Video	4 MHz
Line Difference	3 MHz
Field Difference	2 MHz

and horizontal resolution. As such, it should be sent intact without any differential processing. Other test signals should be devised to measure vertical color resolution in systems, such as proposed here, which limit the transmitted vertical color resolution to that actually required by pictures.

IV. APPLICATION TO A 20-MHz BANDWIDTH MICROWAVE RADIO CHANNEL

4.1 Channel characteristics

Figure 6 shows a typical long-haul, microwave, television transmission system. Preemphasis of high frequencies gives some protection against nonlinearities. Figure 7 gives the preemphasis characteristic; de-emphasis is the inverse of preemphasis. The FMT produces 32.0-MHz deviation per volt input. The gain a is chosen so that a nominal deviation of about 2 MHz results from a one-volt peak-to-peak color bar video signal. However, sustained deviations of 3.2 MHz (found by computer search) can occur for certain input signals, and transient deviations as high as 3.7 MHz (preemphasized 2T pulse)[1] are possible. Ac coupling occurs before the FMT. However, this has little effect on possible peak deviations since its passband extends well below most frequencies of interest.

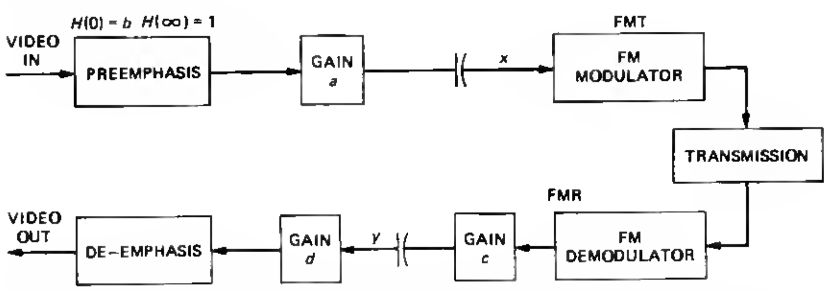

Fig. 6—A typical long-haul television transmission system. $a = -12.6$ dB, $b = -13.3$ dB, $c = 16$ dB, $d = -3.4$ dB, $acd = 1$. De-emphasis is the inverse of preemphasis. The FM modulator produces 32.0-MHz deviation per volt input.

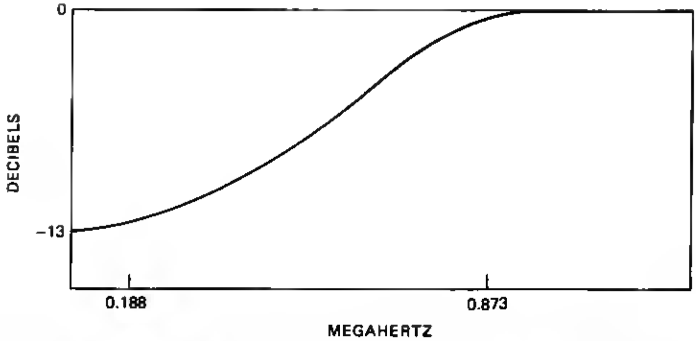

Fig. 7—Preemphasis characteristic. $H(s) = (\omega_1 + s)/(\omega_2 + s)$, where $f_1 = 199$ kHz, $f_2 = 873$ kHz, and $H(0) = b = -13.3$ dB. Preemphasis of high frequencies gives some protection against nonlinearities and improves ANR in colored areas of the picture.

Nonlinearities vary considerably between channels. However, one such microwave radio system (4000-mile terrestrial, 4GHz) was measured, and Appendix B estimates video degradation for a standard VITS. Differential gain is estimated at 8.5 IRE* (13 IRE allowed), and worst case differential phase is estimated at 9.7° (5° allowed). These results indicate that the measured channel was marginal and, therefore, simulation results based on these measurements should be on the conservative side.

4.2 Time-frequency multiplexing

We wish to transmit a baseband signal of 4.2-MHz bandwidth and a single-sideband modulated differential signal in the positive frequency band as shown in Fig. 3. Figure 8 shows a configuration for doing so. It is similar to Fig. 6 except that a 20-MHz bandpass filter following the FM modulator is shown explicitly, and the preemphasis is changed to admit the possibility of attenuating the higher frequencies of the baseband signal somewhat to obtain satisfactory performance.

A line differential signal, which requires 3-MHz bandwidth, is probably not suitable for this application. However, a field differential signal, which requires only 2 MHz of bandwidth, could fit into the 5 to 7-MHz band if sharp cutoff filters were used. Hereafter, results will apply to a system using field differential signals.

The differential signal has no dc component and is small most of the time with ordinary pictures, occasionally reaching values of $\pm 0.36M$.†

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* 1 volt = 140 IRE.

† Measured by computer simulation using SMPTE slides. Electronically generated tests signals can be concocted, however, which produce differential signal values as large as $\pm M$. Here, $M = 0.714$ volt (see Fig. 9).

Carson's rule states that the sum of the total message bandwidth and the FM frequency deviation should not exceed one-half the available channel bandwidth. Since the channel bandwidth is 20 MHz, and the proposed message bandwidth is 7 MHz, the FM deviation should not exceed 3 MHz. Unfortunately, as we have seen, the system in its present form already has deviations larger than 3 MHz, even without the addition of a modulated differential signal. However, these deviations are often short lived and occur at sharp edges in the picture where considerable distortion can be withstood. Still, some signal attenuation is obviously necessary to accommodate a 7-MHz message bandwidth. However, it may not be necessary to attenuate the signals so much that the deviation is always below 3 MHz. If the FMT is followed by a reasonably good 20-MHz bandpass filter, then some of the responsibility for maintaining bandwidth might be shifted to it, at the cost of some picture degradation. This degradation due to the bandpass filter will only occur if the preemphasized baseband signal is large, while at the same time the differential signal is large. In this

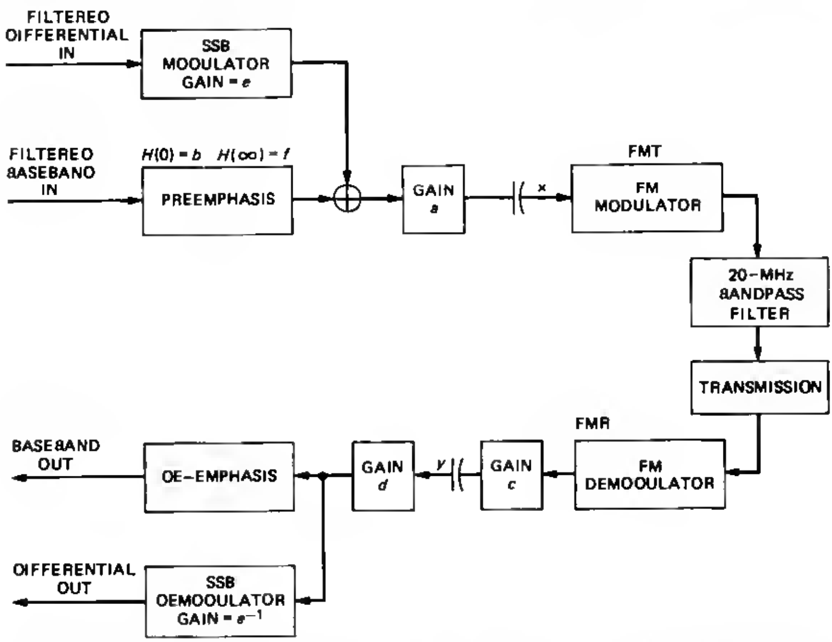

Fig. 8—Configuration for transmitting a single-sideband modulated differential signal. $a = -12.6$ dB, $b = -13.3$ dB, $c = 16$ dB, $d = -3.4$ dB, $acd = 1$. It is similar to Fig. 6 except that a 20-MHz bandpass filter following the FM modulator is shown explicitly, and the preemphasis is changed to admit the possibility of attenuating the higher frequencies of the baseband signal somewhat to obtain satisfactory performance. Gain e causes attenuation of the modulated differential signal (including the effect of single-sideband modulation), and gain f causes attenuation of the high frequencies of the baseband signal. $e = f \approx -3$ dB gives just visible distortion.

case, and in this case only, the bandpass filter will attenuate the modulated differential signal, thus causing distortion. This distortion should be relatively invisible, however, due to its proximity to sharp brightness transitions in the picture. In Fig. 8, gain e causes attenuation of the modulated differential signal (including the effect of single-sideband modulation), gain f causes attenuation of the high frequencies of the baseband signal. Suitable values for e and f must be determined experimentally. However, some guidelines can be obtained intuitively as follows.

If a constant peak-white baseband value of $M = 0.714$ volt (=100 IRE) occurs at the same time as a differential signal peak of $0.36M$, then the input to the FM modulator (assuming worst case DC = 0)* is

$$x = abM + 0.36aM \cos'\omega_D t, \tag{1}$$

where ω_D is the differential signal carrier frequency, and cos′ indicates single-sideband modulation. If the conservative position is taken, of never allowing a deviation above 3 MHz, then from values previously defined, e should be less than −0.4 dB. This is a very modest attenuation.

Let us consider another input signal. A maximum sustained preemphasized baseband signal occurs if the video input is $0.52M + 0.48M \cos\omega_c t$. If this value occurs at the same time as a peak differential signal value of $0.36M$, then again assuming the worst case (DC = 0),

$$x = 0.52abM + 0.48afM \cos\omega_c t + 0.36aeM \cos'\omega_D t. \tag{2}$$

If, in this case, the very conservative approach of 3-MHz maximum deviation is taken, then with an FMT deviation of 32 MHz/volt,

$$0.52abM + 0.48afM + 0.36aeM \leq 0.094 \text{ volt}. \tag{3}$$

If equal values are chosen for e and f,

$$e = f \leq -5.4 \text{ dB}. \tag{4}$$

Although choosing e and f to satisfy Eq. (4) eliminates almost all distortion caused by the 20-MHz bandpass filter, it also leads to a decrease in overall SNR, i.e., fade margin. Since some of the bandpass filter distortion will be invisible, it seems wasteful to use such low values for e and f. Larger values should be feasible with the same subjective picture quality. Thus, another of the reasons for performing the computer simulations was to estimate the range of acceptable values for e and f.

* Average signal level. This is not transmitted in an ac coupled system.

V. COMPUTER SIMULATION OF NONLINEARITIES

Since the transmission channel is not precisely understood and is time varying as well, such simulations can only be expected to give a vague idea of overall performance. Hopefully, however an estimate of worst-case performance can be obtained if a conservative approach is taken. The channel characteristics used in Appendix B and shown in Table II probably represent a marginally acceptable situation insofar as nonlinearities are concerned. Thus, the use of these data in the simulations should give conservative results.

There are several approaches to modeling the nonlinearity, including the rather complicated procedure of numerically carrying out the complete SSB modulation, multiplexing and FM operations, and passing the result through the nonlinearity. However, the simulations to be described below were all carried out at baseband for simplicity and since intermodulation crosstalk was the main item of interest.

The basic approach taken in evaluating the effects of nonlinearities was to introduce degradations into the NTSC composite signal and see if they were visible in the resulting reproduced picture. Two types of degradation were studied: (*i*) intermodulation crosstalk resulting from nonlinearities, and (*ii*) the effect of the 20-MHz bandpass filter following the FM modulator in Fig. 8.

Baseband and differential signals were first computed as in Section III for each of the original SMPTE pictures. Intermodulation crosstalk terms were then estimated as described in Appendix C. Some of the terms degrade the baseband signal, and some degrade the differential signal.

The effect of the 20-MHz bandpass filter was approximated as follows: If at certain times the sum of the baseband and differential signal magnitudes at the input to the FMT was large enough to produce a 3-MHz deviation (again assuming worst case dc = 0), then the magnitude of the differential signal was reduced until the deviation fell below 3 MHz. If the baseband signal by itself was large enough to

Table II—Results of two-tone measurements on a typical long-haul network. $A = B = -26$ dBv, $c = 16$ dB. From the measured harmonic outputs, the coefficients u and v of the third-order model can be obtained.

α (MHz)	β (MHz)	$ucAB$ FMR average dBv at $\beta + \alpha$ and $\beta - \alpha$	$\frac{3}{4}vA^2Bc$ FMR average dBv at $\beta + 2\alpha$ and $\beta - 2\alpha$	u	v
0.1	6.55	−41	−49	0.56	5.96
0.1	3.55	−35	−46.5	1.12	7.94
0.1	1.55	−36	−39	1.00	18.84

produce a 3-MHz deviation, then the differential signal was set to zero.*

In the simulations, gain factors a through f were all incorporated as shown in Fig. 8; factors a through d were fixed having the values shown, while e and f were made variable. The nonlinearity coefficients were taken from Table II, depending on the applicable frequency range.

Results were generally encouraging, considering the pessimistic and conservative assumptions built into the simulations. Using attenuation factors $e = f = 0.5$ resulted in no visible degradation, as expected from eq. (4). Larger values in the range 0.7 to 0.8 (−3 to −2 dB) gave rise to barely visible color shading in a few areas such as the boundary between red lipstick and white teeth. Otherwise, the effects were invisible.

VI. CONCLUSION

Color pictures were processed via computer simulation to test the feasiblity of using time-frequency multiplexing to transmit two broadcast-quality color television signals over a 20-MHz bandwidth microwave radio channel. We considered bandwidth allocation and nonlinearities in detail. At every turn in the study, conservative assumptions were made. The displayed pictures produced full-color bandwidth even though very few home or studio monitors in the U.S. are capable of doing so. Nonlinearity parameters which were used in the simulations were obtained from an, at best, marginal channel. Worst-case theoretical nonlinearity impairments were assumed. And, finally, for picture quality assessment an A–B comparison on the same monitor was used, which is one of the most stringent tests known.

Results were rather encouraging. There appeared to be enough bandwidth to accomplish the transmission. With a modest 3-dB reduction of signal level, simulated transmission defects were practically invisible.

Utilization of field differential signals requires field memories at the transmitter and receiver. Although they are expensive at present, their cost is dropping. Moreover, if synchronization of two pictures is required at the transmitter, then the memory of the synchronizer can be used.

Audio transmission has not been considered here. One might be tempted to place audio carriers in the vacant frequency band between the baseband and differential signals. However, this would require a rather linear channel, which is the very problem that the techniques

* An automatic gain control could, in fact, be implemented in this way to guarantee deviation less than 3 MHz.

of this paper are meant to avoid. A better solution would be to transmit the audio digitally, either in the horizontal sync pulse[2] or during vertical blanking.

APPENDIX A

Computer Simulation of NTSC Color Multiplexing and Demultiplexing

The computer simulation facility has a picture resolution of 256 × 256 pels per frame. At a frame rate of 30 Hz, 15 lines vertical blanking, 17 percent horizontal blanking, this implies a sampling rate of ≈2.51 MHz. Thus, video signal bandwidth of the computer is conservatively estimated to be about 1 MHz. Since U. S. broadcast rate NTSC color television signals have a nominal bandwidth 4.2 MHz, scaling by a factor of approximately four is required to do meaningful simulations.

The NTSC broadcast color subcarrier frequency is approximately 3.58 MHz. Thus, for the simulations, a value around 0.89 MHz should be used. In addition, the simulation color subcarrier frequency *must* be an odd multiple of half the line scanning frequency of the simulation system ($f_H \approx 8.1$ kHz). Thus, a convenient value for color carrier frequency that was chosen for the simulations was $f_c = 108.5 f_H \approx 0.88$ MHz.

The simulated sampling rate of $4f_c$ equals $434 f_H$ or 434 pels per line. Assuming about 17 percent horizontal blanking, we have 358 pels in the active area of the picture (which corresponds to the 256 pels of the display). To convert the original red, blue, and green (RBG) signals having 256 pels per line to the desired 358 pels per line, linear interpolation was used. This sampling rate conversion is suboptimal in that a frequency rolloff of 4.7 dB at 1 MHz is introduced. The extra effort required to design optimal digital filters for sampling rate conversion did not seem worthwhile at this time. After processing, the conversion back to 256 pels per line was again performed via linear interpolation.

From the RBG signals, the full bandwidth Y, I, Q signals* were obtained by the standard matrix relation[3]

$$\begin{bmatrix} Y \\ I \\ Q \end{bmatrix} = \begin{bmatrix} 0.3 & 0.11 & 0.59 \\ 0.6 & -0.32 & -0.28 \\ 0.21 & 0.31 & -0.52 \end{bmatrix} \begin{bmatrix} R \\ B \\ G \end{bmatrix} \qquad (5)$$

The I and Q signals were then bandlimited to ≈0.375 MHz and 0.125 MHz, respectively (0.25 × NTSC values) using symmetrical finite-impulse-response digital filters to avoid phase distortion. Filters were chosen to meet NTSC specifications, yet to be short enough to avoid excessive ringing.

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* Y = luminance, I and Q = chrominance signals.

Samples at $4f_c$ were assumed to occur at the peaks and zeros of the I and Q carrier sinusoids. The chrominance (CH) pels could then be formed simply from

$$\pm CH(N) = I(N)\cos(N\pi/2) + Q(N)\sin(N\pi/2), \qquad N = 1, \cdots, 358, \quad (6)$$

the plus and minus signs being taken on alternate lines of the field. The chrominance and luminance were then added together and filtered to a bandwidth of 1 MHz, again with a FIR digital filter to avoid phase distortion. The result is the composite NTSC color signal.

Recovery of the R, B, G signals from the composite signal is not quite so straightforward if full bandwidth for the luminance and chrominance is to be maintained. Comb filtering of three successive lines ($-0.25L_1 + 0.5L_2 - 0.25L_3$) followed by bandpass filtering (center frequency f_c) was used to recover the chrominance signal for the center line. Simple subtraction of the chrominance from the composite signal yielded the luminance pels.

Recovery of I and Q from the chrominance started with a quadrature AM demodulation,

$$I(N) = \pm CH(N)\cos(N\pi/2),$$
$$Q(N) = \pm CH(N)\sin(N\pi/2), \quad (7)$$

the plus and minus signs being taken on alternate lines. The required low-pass filtering was implemented simply by replacing zero samples by an average of their neighbors.

At this point, the I signal required equalization since it lost part of its upper sideband during the 1-MHz low-pass filtering of the composite signal. The frequency range from 0.125 MHz to 0.375 MHz must be increased by 6 dB. This was accomplished, again, by a FIR digital filter. Having obtained Y, I, and Q, the R, B, G signals were produced by inverting the matrix equation (5) above.

APPENDIX B

Effects of Nonlinearities on the Vertical Interval Test Signal (VITS)

In Fig. 6, nonlinearities are often modeled as third-order polynomials,[4]

$$y = f(x) = c(x - ux^2 - vx^3). \quad (8)$$

This model is reasonable for transmission systems with "small" nonlinearities. The nonlinearity coefficients u and v are often estimated by sending the two-tone signal

$$x = A\cos\alpha t + B\cos\beta t, \quad (9)$$

in which case y/c consists of a number of sum and difference frequencies as follows:[4]

$$\text{DC: } u(A^2 + B^2)/2 \quad \text{(deleted at FMR)} \tag{10}$$

$$\text{First Order: } A\cos\alpha t + B\cos\beta t + \text{very small terms,} \tag{11}$$

$$\begin{aligned}\text{Second Order: } & -\tfrac{1}{2}u(A^2\cos 2\alpha t + B^2\cos 2\beta t) \\ & -uAB[\cos(\alpha+\beta)t + \cos(\alpha-\beta)t],\end{aligned} \tag{12}$$

$$\begin{aligned}\text{Third Order: } & -\tfrac{1}{4}v(A^3\cos 3\alpha t + B^3\cos 3\beta t) \\ & -\tfrac{3}{4}vA^2B[\cos(2\alpha+\beta)t + \cos(2\alpha-\beta)t] \\ & -\tfrac{3}{4}vB^2A[\cos(2\beta+\alpha)t + \cos(2\beta-\alpha)t].\end{aligned} \tag{13}$$

In 1974, two-tone measurements were made of a typical long-haul network. From these measurements the coefficients u and v can be estimated. Results are frequency dependent and are shown in Table II for $A = B = -26$ dBv,* and c = 16 dB.

These data can be used to estimate degradations which would occur in the transmission of a vertical interval test signal (VITS). Let M = 100 IRE = 0.714 volt be the blanking-to-peak magnitude of the video signal as shown in Fig. 9. Values of gains a, b, and c are given in Fig. 6.

Differential gain (Ref. 1, Section 3.13)

$$\text{dc} = 0,$$

$$\text{Peak input} = M(0.9 + 0.2\cos\omega_c t). \tag{14}$$

Thus,

$$\begin{aligned} x &= aM(0.9b + 0.2\cos\omega_c t) \\ &\triangleq A + B\cos\omega_c t, \end{aligned} \tag{15}$$

where A is the low-frequency magnitude, and B is the color-subcarrier magnitude. From eqs. (11), (12), and (13) with $\alpha \approx 0$ and $\beta = \omega_c$, the color-subcarrier terms in y/c (when phases align,[5] which is worst case) are

$$B - 2uAB - \tfrac{3}{2}vA^2B. \tag{16}$$

Thus, at ω_c the signal and distortion components in y are, respectively,

$$\begin{aligned} R &= Bc = 0.2aMc, \\ \Delta R &= -2uABc - \tfrac{3}{2}vA^2Bc. \end{aligned} \tag{17}$$

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* Zero to peak.

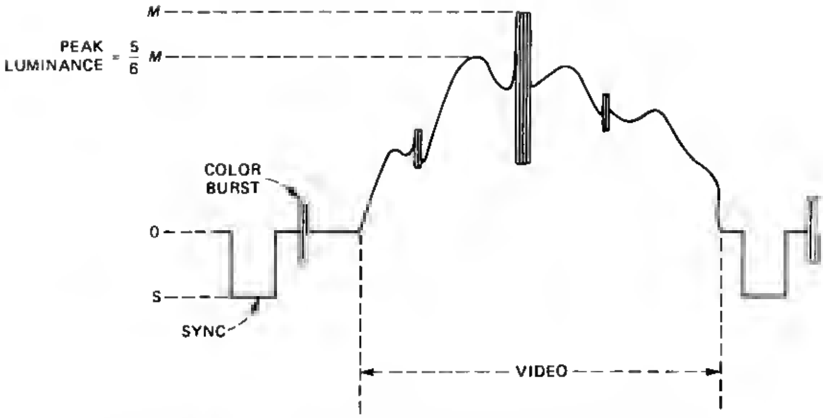

Fig. 9—Composite video signal, including sync, color burst and, video. $S = -0.286$ volt. According to specifications, M can be as large as 0.857 volt. However, to produce, the normally used one-volt peak-to-peak composite signal, $M = 0.714$ volt (=100 IRE).

De-emphasis has no effect at this frequency. Thus, according to the test procedure we should divide the above equations by $0.2ac$ to bring the signal component up to M. Then, peak differential gain is given by

$$\Delta G = \frac{-2uABc - \tfrac{3}{2}vA^2Bc}{0.2ac}. \tag{18}$$

Using values defined previously and $u = 1.12$, $v = 7.94$ from Table II,

$$\Delta G = -0.06 \text{ volts} = -8.5 \text{ IRE}, \tag{19}$$

which is well within the 15 IRE allowed.[1]

Differantial phasa (Ref. 1, Saction 3.14)

The same input signal is used as in the differential gain test. Using the same definitions for signal R and distortion ΔR, worst-case differential phase occurs when these two components are exactly 90° out of phase.[5] In this case, peak-to-peak differential phase is

$$\Delta\phi = 2 \tan^{-1} \frac{\Delta R}{R} = 2 \tan^{-1} A\left(2u + \frac{3}{2} vA\right), \tag{20}$$

which for the values defined previously evaluates to

$$\Delta\phi = 9.7°,$$

which is somewhat outside the 5° allowed. However, this is a worst-case maximum and might not occur in practice.

Concluaion

From these estimates of distortion one concludes that the channel measured was not of extremely high quality and may have been only

of marginal quality. Thus, simulation results using those measurements should be on the conservative side.

APPENDIX C

Important Intermodulation Crosstalk Terms

First consider crosstalk between baseband color-subcarrier frequency and the differential signal. Let BB_c be the input baseband chrominance component (frequencies $= \omega_c \approx 3.58$ MHz) and let D be the input differential signal (carrier frequency $\omega_D \approx 7$ MHz). Then, from Fig. 8,

$$x = af \cdot BB_c \cos\omega_c t + aeD \cos\omega_D t$$
$$\triangleq A_1 \cos\omega_c t + B \cos\omega_D t, \tag{21}$$

as in eq. (9). Then, from eqs. (11) to (13) the corresponding in-band distortion components in y are

$$-\tfrac{1}{2}ucA_1^2 \cos 2\omega_c t - uA_1B \cos(\omega_D - \omega_c)t - \tfrac{3}{4}vcA_1^2B \cos(\omega_D - 2\omega_c)t. \tag{22}$$

Combining eqs. (21) and (22) yields estimates of interfrequency crosstalk caused by the above frequency components. In the computer simulations BB_c was obtained by first bandpass filtering the composite signal to retain only components from 2 to 4 MHz. Tbe result was then shifted down in frequency by 2 MHz to obtain the signal BB_c used in eqs. (21) and (22) to compute distortion.

Similar crosstalk occurs between baseband low frequencies and the modulated differential signal. However, because of preemphasis, A in eqs. (11) to (13) is small, making this distortion negligible.

Crosstalk between baseband midfrequencies, e.g., ω_M around 1 MHz, and the modulated differential signal can be estimated similarly. Let BB_M be the input baseband component in the midrange frequencies ($\omega_M \approx 1$ MHz), and define D as above. Then from Fig. 8,

$$x = af \cdot BB_M \cos\omega_M t + aeD \cos\omega_D t$$
$$\triangleq A_2 \cos\omega_M t + B \cos\omega_D t. \tag{23}$$

From eqs. (11) to (13) the in-band distortion components in y are

$$-\tfrac{1}{2}ucA_2^2 \cos 2\omega_M t - ucA_2B \cos(\omega_D - \omega_M)t$$
$$-\tfrac{1}{4}cvA_2^3 \cos 3\omega_M t - \tfrac{3}{4}vcA_2^2B \cos(\omega_D - 2\omega_M)t. \tag{24}$$

In the computer simulations, BB_M was obtained by low-pass filtering (2-MHz cutoff) the composite signal.

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